Iterative Equalization Enhanced High Data Rate in Wireless Communication Systems

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***Abstract*-There has been a demand for high data rate for wireless communication system devices, mainly mobile multimedia applications. This paper investigates the suitability of Turbo Equalization as a means of achieving low bit error rate in the future high data communication systems. Turbo equalization is an approach to coded data transmission over channels with Inter-symbol interference (ISI) which can yield additional improvement in bit error rate. The paper demonstrates that at higher modulation scheme using iterative equalization method, high date rate can be achieved. The performance evaluation shows that turbo equalization is beneficial for higher modulation and thereby increase data rate with a reasonable complexity.**

***Keywords*-Equalization, Decoding, Iterative methods, Turbo equalization**

I. INTRODUCTION

Data transmission over IS1 channels is a classical problem in communication systems. Conventional approaches implement an equalizer to remove ISI or use MAP or maximum likelihood (ML) detection. Data reliability can be enhanced using coding, when the data is encoded in the transmitter prior to transmission. For reasons of complexity, the receiver then typically performs separate equalization and decoding of the data. Significant performance gains *can* be achieved through joint equalization and decoding at the cost of added complexity.

A recent approach that significantly reduces the complexity of joint equalization and decoding is the so called “turbo equalization” algorithm, where MAP/ML detection and decoding are performed iteratively on the same set of received data [1, 2]. It has recently been shown that passing soft information, the use of interleaving, and the controlled feedback of soft information are essential requirements to achieve performance gains with an iterative system [3]. Various algorithms similar to Turbo Equalization have been proposed to overcome the complexity of the MAP/ML algorithms, especially in the detector, where complexity is exponential in the channel delay spread [4, 5]. Since the initial proposal of ‘Turbo Codes’ by Berrou et *al* in 1993 [6], the iterative principle has been extended to encompass single carrier equalization techniques. This allows single carrier systems to combine the operations of equalization and channel coding to operate in a wideband channel with performance that could not previously be achieved with traditional equalization and forward error correcting (FEC) techniques [7]. Iterative equalization techniques have been shown to give excellent error rate performance for both fixed and fast fading channels [8].

This paper investigates the suitability of including Turbo equalization in the receiver to combat ISI. BER results are presented for BPSK, QPSK, 8PSK and 16QAM modulation schemes in a slow Raleigh fading channel scenarios. The channel models are characterized in terms of normalized delay spread, which allows comparison independent of system symbol rate. This paper is organized as follows. Section II describes the transmission system model for communication system; the principle of Turbo Equalization is depicted in detail in section III. Section IV describes the receiver algorithms MAP (Maximum A Posteriori). Data rate performance analysis in section V, Performance results and conclusion in section VI and VII respectively.

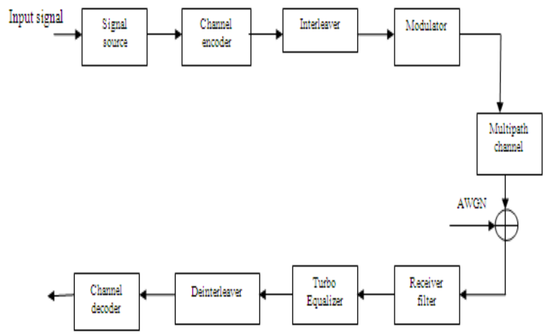

Figure 1 Communication System Model

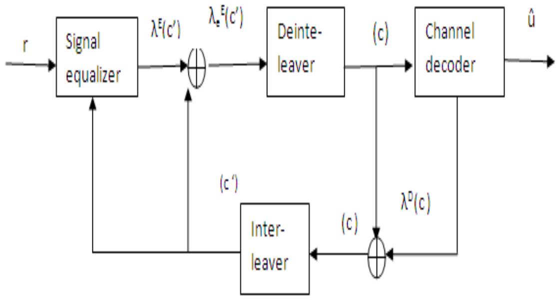

Figure 2 The Iterative Structure

II. SYSTEM MODEL

In this paper, we used the digital communication platform, which is modelled in Figure 1. In the transmitter side, a block of data bits **u** is protected by convolutional encoder and interleaved to overcome fast fading phenomenon. The encoded bits $a = (a_0, a_1,..a_{3k-1})^T$ are modulated. Those symbols are denoted by $b = (b_0,b_1,\ldots b_{k-1})^T$ .The modulated signal is transmitted over a frequency selective fading channel. In this paper we assume block fading channel characteristics; hence the channel is time-invariant during one transmission burst. Thermal noise at the receiver is modelled as additive white Gaussian noise (AWGN). The received signal **r** that is sampled at the symbol rate can be given by the equation

$$r = Ah + w, \qquad (1)$$

where **A** is the matrix containing symbols.

The channel impulse response is described by the vector **h** = $(h_0, h_1 \dots h_L)^T$, which consists of symbol spaced complex-valued channel taps. The white noise samples are denoted by **w**; the noise variance is $\sigma^2 = N_0/2$.

III. PRINCIPAL OF TURBO EQUALIZATION

Figure 2, shows the iterative equalization receiver structure used in this study. Both the equalizer and the decoder employ the optimal symbol by symbol Maximum A-Posteriori (MAP) soft input soft output (SISO) algorithm [13]. Soft input symbols are fed into the decoder from a sampled receive filter stream ***r(t)*** and bit-wise hard decisions are produced as the final output. It is possible to equalize and decode in an iterative manner that is similar to turbo decoding. The equalizer provides soft outputs, i.e., reliability information on the coded bits for the channel decoder. The soft information on the bit *c*k is usually given as a log-likelihood ratio (LLR) or *L*-value

$$\lambda^E(\mathbf{c}_k) = \log \frac{P(ck=+1|r)}{P(ck=-1|r)} \quad (2)$$

This is the ratio between the conditional bit probabilities in the logarithmic domain. These *L*-values are deinterleaved and given for the channel decoder, which uses them to recover the information bits **u**.At the first iteration round there is no feedback information from the channel decoder available, so the equalizer calculates the *L*-values $\lambda^E(\mathbf{c}')$ as given by (4) that are just based on the received samples **r** from the channel. The *L*-values are deinterleaved to break consecutive bits far apart and thus giving the channel decoder independent input values. The interleaving is an essential part in the iterative receiver scheme, since the extra information on an individual data bit is due to the different neighboring bits in the detection and decoding processes.The soft values $\lambda^E(\mathbf{c}')$ are provided for the SISO channel decoder. The decoder has to calculate new *L*values $\lambda^D(\mathbf{c}')$ for the coded bits **c**, since they are needed in the feedback branch to the equalizer. Therefore we need to use the more complex SISO decoder instead of the conventional hard output decoder. The equalizer is able to produce the *L*-values $\lambda^E(\mathbf{c}')$ based on the received samples from the channel, so that information should not be repeated in the feedback. Hence, the feedback only contains the extra information that is obtained from the surrounding bits in the channel decoding. The input *L*-values and the obtained extra information are called intrinsic and extrinsic information, respectively. The extrinsic information from the channel decoder is given as [2, 3]

$$\lambda_e^D c_k = \lambda^D(\mathbf{c}) - \lambda_e^E(\mathbf{c}_k), \quad (3)$$

where $\lambda_e^E(\mathbf{c}_k)$ denotes the extrinsic information from the equalizer. The turbo equalization technique is based on the utilization of this extrinsic information at the next iteration round [2]. So it is passed through the interleaver to the equalizer as *a priori* information on the bit reliabilities. By exploiting this side information in the detection, more reliable decisions are achieved. Also in the equalizer output the extrinsic information $\lambda_e^E(\mathbf{c}_k)$ is extracted from the output as follows

$$\lambda_e^E c_k = \lambda^E(\mathbf{c}) - \lambda_e^D(\mathbf{c}_k), \quad (4)$$

This equalizer information is again used in the SISO decoder to produce new soft outputs and furthermore, the new extrinsic information according to (3). As soon as this feedback information becomes available, the new iteration round can be started. The number of iterations may depend on the processing power available or the achieved performance improvement. At the final stage, there is no need for the SISO decoder, since only hard decisions û on the information bits are needed. The Turbo equalization receiver is able to improve the performance, but at the cost of higher complexity. The main burden is the complex SISO decoder, especially due to the coding schemes that are based on the constraint length of 4. Also, as the equalization and decoding are performed several times, the receiver complexity grows respectively.

IV. RECEIVER ALGORITHM

The objective of the MAP algorithm is to minimize the bit error probability by estimating a posteriori probabilities (APP) of states and transitions of the Markov source from the received signal sequence. The MAP algorithm introduced by Chang and Hancock [9] uses the information of the whole received sequence to estimate a single bit probability. In other words, the MAP algorithm selects the bit û ={-1, +1} at time instant *k*, which maximizes the following APP

$$\hat{u} = \arg\max\,[p\,(u|\mathbf{c})]. \quad (5)$$

The optimum MAP algorithm saves multiplicative transition probabilities in the trellis, which is computationally difficult. Therefore in practical implementations the path metric is calculated in the log-domain, which enables cumulative calculations [10, 11]. Since MAP requires both forward and backward recursions, it is around two times more complex than the soft output Viterbi algorithm (SOVA) [11]. The main advantage of the MAP algorithm is more reliable soft information, since it is optimized for symbol wise decoding. This is why MAP is very suitable for the SISO decoding algorithm in the Turbo equalization scheme. The BCJR-log-MAP (BCJR stands for Bahl, Cocke, Jelinek and Raviv) provides the APP information for each bit as the *L*-value according to (2). The state probability for trellis state *s* at time *k* is denoted by

$$\alpha_k(s) = p(s, r_{j\le k}) \quad (6)$$

in the forward direction and by

$$\beta_k(s) = p(r_{j>k}|s) \quad (7)$$

in the backward direction. The transition probability between states *s'* and *s* is given in log-domain as

$$\ln\gamma_k(s',s) = \sum_{i=0}^{N} \frac{1}{2}\lambda_e^E(\mathbf{c}_k)c_{k,i}, \quad (8)$$

where $c_{k,i}$ is the *i*th code bit for the information bit *uk* and the coding rate is 1/*N*.

Then the decoder output is given as [11, 12]

$$\lambda(\hat{u}) = \ln\sum_{(s',s),\ u_k=+1} e^{\ln\alpha_{k-1}(s')+\ln\gamma_k(s',s)+\ln\beta k(s)} - \ln\sum_{(s',s),\ u_k=-1} e^{\ln\alpha_{k-1}(s')+\ln\gamma_k(s',s)+\ln\beta k(s)} \quad (9)$$

and the state probabilities can be computed recursively

$$\ln\alpha_k(s) = \ln\sum_{s'} e^{\ln\gamma k(s',s)+\ln\alpha_{k-1}(s')} \quad (10)$$

$$\ln\beta_k(s') = \ln\sum_{s} e^{\ln\gamma_{k+1}(s',s)+\ln\beta_{k+1}(s)} \quad (11)$$

Finally, the soft information on the code bits $\lambda(c_k)$ (needed for the feedback) is obtained by re-encoding the achieved output (19)

V. DATA RATE PERFORMANCE ANALYSIS

The data rate R_D against SNR_{dB} can be calculated from the packet error rate as follows:

$$R_D= (Rs*K)*(R)*(1\text{-}PER) \quad (12)$$

where K is the bandwidth efficiency, R is the packet code rate. It is assumed that no packets are discarded due to header errors. The PER is purely a function of the payload error rate. This quantative measure allows us to compare each of the modulation modes and channel scenarios in terms of the data rates versus SNR.

$$SNR_{dB}=P_R–N \quad (13)$$

In order to examine the data rates versus signal to noise ratio, a link budget must be established [15, 16]. To calculate the SNR, given by (12) at the receiver, we must consider expressions for the noise power N and also the receiver power P_R in terms of the transmit power P_T and the path loss P_L:

$$N =10\log (K_BT) + 10 \log_{10} (R_S * (1+\alpha)) + NF \quad (14)$$

Where K_B is Boltzmann's constant (1.38 x 10^{-23}W-s/K) and T is the noise temperature in Kelvin. NF is the total noise power in signal bandwidth and:

$$P_R = P_T + P_L$$

$$P_L= G_T+G_R+20\log10(\frac{\lambda}{4\pi})\text{-} n*10\log_{10} (d) \quad (15)$$

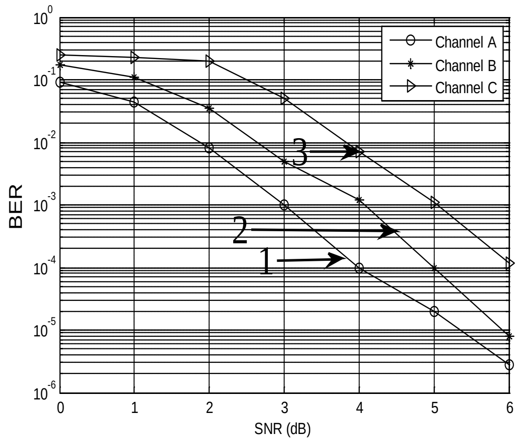

Figure 3 The BER of the three channel model Comparison as a function of SNR1) Channel A 2) Channel B 3) Channel C

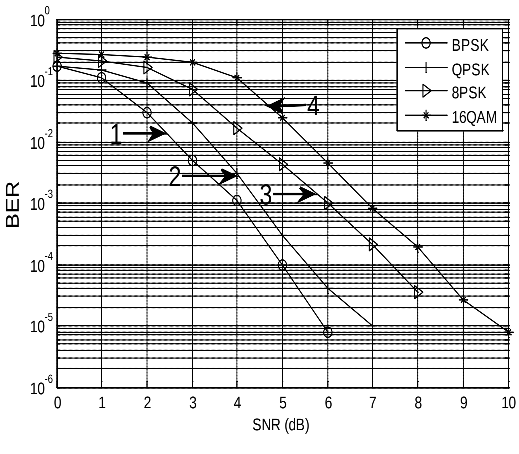

Figure 4 BER using BPSK, QPSK, 8PSK and 16QAM1) BPSK 2) QPSK 3) 8PSK 4) 16QAM

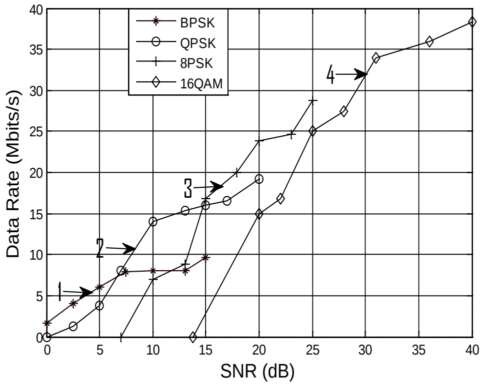

Figure 5 Data rate for channel B using BPSK, QPSK, 8PSK,and 16QAM as a function of SNR.1) BPSK 2) QPSK 3) 8PSK 4) 16QAM

where G_T and G_R are the gains of the transmit and receiver antennas respectively relative to an isotropic source, A is the wavelength of the carrier frequency and d the separation between the antennas. n is an empirical constant, the path loss exponent, which for line of sight (LOS) is *2* and greater than *2* for non-LOS conditions. We assume a path loss exponent of 6, which represents a worst case scenario. The maximum actual data rates achievable for the different modulation modes are shown in Figure 5 calculated from (12) assuming a symbol rate of 20MHz.

VI. PERFORMANCE RESULTS

The BER performance of an iterative equalizer is dependent upon the channel profile, the modulation scheme, the encoder constraint length and the size of the interleaver. In this paper the encoder constraint length and the size of the interleaver are fixed. The result demonstrates the effect of changing the channel delay spread with fixed modulation scheme. And also for a fixed delay spread as the modulation scheme changes. The channel code rate is ½ recursive systematic convolutional with the generator polynomials [7 5] in octal form. The frame size is 2^{14} bits while the filter length is 15. Figure 3 shows the BER performance as a function of SNR, when BPSK modulation is used with the three channel models while Figure 4 shows the performance of the BER as a function of SNR for using BPSK, QPSK, 8PSK, and 16QAM using channel model B only and decoded with MAP algorithm [8]. If we take a target of 10^{-4}, then to achieve this target BPSK mode required 5dB, QPSK 5.1dB, 8PSK 7.4dB, 16QAM 8.3dB respectively. As the modulation order increases, the iterative gain increases. Figure 5 shows the date rate achieved for BPSK, QPSK, 8PSK, and 16QAM reaching their maximum data rate at 9.6, 19.2, 28.8, and 38.4 Mbits/s respectively. The bit rate increased even further by introducing higher order modulation such as 16QAM.

VII. CONCLUSION

The result for higher modulation scheme shows that as the modulation order increases, a higher SNR is required to obtain the same BER performance as a lower order. The iterative gain is greater for higher modulation orders. However, there is a trade off, between the iterative gain for higher modulation orders and the complexity of the receiver. The complexity at the receiver is dominated by the complexity of the MAP equalizer. The number of states in the equalizer trellis is dependent upon both the modulation order and the memory of the channel. In the decoder the trellis size is only dependent on

the constraint length of the code. The BER performance improves as the delay spread in the channel increases.

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